Space-time front-ends for RAKE receivers in the FDD mode of UTRA[1]

Josep Vidal, Margarita Cabrera, Adrian Agustín
Signal Theory and Communications Department, Universitat Politècnica de Catalunya
Barcelona, Spain
{pepe,marga}@gps.tsc.upc.es

Abstract

A unified and general vision of different space-time processors is presented. Many popular receivers can be accomodated, like V-RAKE receivers, weighted V-RAKE, or spatial narrowband beamforming. By making appropriate assumptions on the space/time characteristic of the interference it is possible to enhance the performance of the receiver through spatial/temporal pre-processors. These receivers will be tested in the FDD mode of UTRA.

I. Introduction

The advent of the 3rd generation of mobile communications systems has been accompanied by the recognition of the large increase in system capacity that can be obtained from the use of adaptive antenna arrays. Care has been taken in the definition of the standard to include capabilities for space-time processing of the signals incoming and radiated from the base stations. Section II reviews the main characteristics of the European proposed air interface (UMTS). Section III presents the signal model. In section IV the different space-time receivers are presented in a unified vision. It will be seen that the use of multiple antennas and the temporal correlation diversity of multiple users allow additional degrees of freedom for cancellation of multiple access interference. Only single user approaches will be introduced although some ideas are easily extendable to the multiuser case. All these receivers need a reliable estimate of the correlation properties of the interferences, so different options are presented in section V. Performances are compared in terms of probability of error in section VI.

II. UTRA FDD uplink air interface

The uplink physical channels of the FDD mode of UTRA are as follows: each user generates at least one Dedicated Physical Data Channel (DPDCH) along with a single Dedicated Physical Control Channel (DPCCH). Each channel is spreaded with a different OVSF orthogonal code at a chip rate of 4.096 Mchip/s and then in-phase and quadrature multiplexed in a QPSK modulation. The existence of OVSF codes of different length allows the presence of different bit-rate users in the same cell. These complex symbols are scrambled by a mobile-station specific scrambling code using either codes from the Very Large Kasami set or the Gold sequence of length 40960 (figure 1 illustrates the signal generation scheme [ETSI-UTRA]).

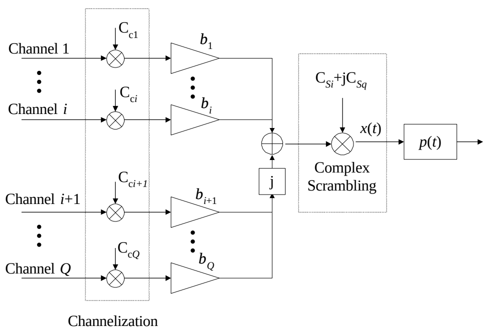


Figure 1. Modulated signal generation

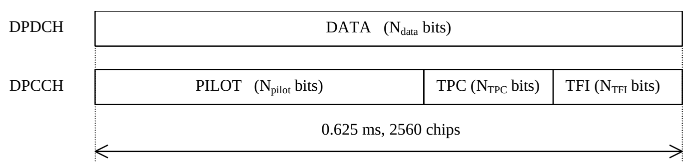


Figure 2. Slot structure for the uplink in the FDD mode

The Kasami codes are intended for base stations featuring multiuser detection, while in the Gold codes assure the separation of users on the grounds of good correlation properties of the codes and single user detectors have to be used. For a review of multiuser detectors for array observations the reader may resort to [Jung][Verdu].

The UTRA air interface is designed to achieve a full 1:1 reuse, which implies no loss of capacity due to frequency planning. Since all users access the channel asynchronously, intercell and intracell users separation

[1] This work has been done thanks to the AECI, EC (IST-199-10322, SATURN project), CICyT of Spain (TIC98-0412, TIC98-0703, TIC99-0849) and CIRIT of Catalunya (1998SGR-00081).

relies on the good correlation properties of the scrambling codes. Under this premises, all channelization OVSF codes are available to each user so many DPDCH can be set up to obtain the desired bit-rate (as shown in figure 1) with different weighting factors (noted with β). The resulting bit-rate granularity is really high.

Due to the severity of the mobile channel, the dynamic range of received powers in the BS can be as high as 90 dB. In order to avoid loss of efficiency due to the near-far problem, a tight control of the transmission power is introduced: every 0,625 ms, the base station sends power control information through the forward link, and the mobiles update their transmitted power at the same frequency.

The frame structure is also shown in figure 2. Each frame of duration 10 ms is split into 16 slots of duration 0,625 ms, which corresponds to one power-control period. The DPDCH is used to transport data symbols, while the DPCCH is divided in three fields which are used respectively for channel estimation (PILOT), transmission of power control information (TPC) for the downlink and transport-format indicators (TFI). TPC and TFI occupy only the last 10% of the slot duration, so the receiver is able to estimate and track the channel for the first 90%. Usually the channel estimation is made using the pilot chips, and different amplitudes can be assigned to the DPDCH and the DPCCH chanels.

III. Signal model

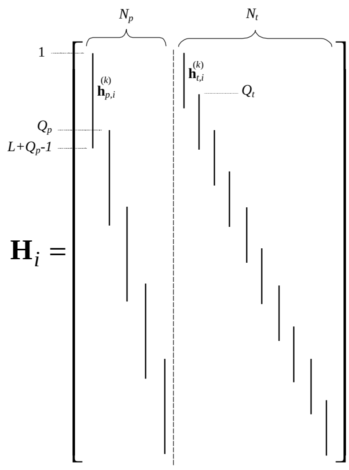
The single-user signal model assumed for the signal received at M sensors after matched filtering and chip-time sampling can be writen in column vector form as:

$$\mathbf{y} = \mathbf{Hd} + \mathbf{w} \tag{1}$$

where each term is defined as:

$$\begin{aligned}
\mathbf{d}^T &= \left[\mathbf{d}_p^T \mathbf{d}_t^T\right] \\
\mathbf{y}^T &= \left[\mathbf{y}_1^T \;\vdots\; \mathbf{y}_2^T \;\vdots\; \cdots \;\vdots\; \mathbf{y}_M^T\right] \\
\mathbf{y}_i^T &= \left[y_i(0) \quad y_i(1) \quad \cdots \quad y_i(N-1)\right]
\end{aligned}$$

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}_1 \\ \hdashline \mathbf{H}_2 \\ \hdashline \vdots \\ \hdashline \mathbf{H}_M \end{bmatrix} \qquad \mathbf{H}_i = \tag{2}$$



$\mathbf{H}$ is the space-time channel of the desired user, $\mathbf{d}$ includes the N_t traffic symbols ($\mathbf{d}_t$) and the N_p pilot symbols ($\mathbf{d}_p$) for this user and $\mathbf{w}$ is the vector accounting for noise plus interferences. N is the number of chips in a single slot, Q_p is the DPCCH spreading factor and Q_t is the DPDCH spreading factor. Matrix $\mathbf{H}_i$ contains the convolution of the impulse response of the channel seen by sensor i (computed at chip time) and the spreading code. The effect of the long scrambling code can be represented by the time variation of the spreading code from symbol to symbol, which is denoted with the superscript (k):

$$\begin{aligned}
h_{i,p}^{(k)}(n) &= h_i(n) * c_p^{(k)}(n) \\
h_{i,t}^{(k)}(n) &= h_i(n) * c_t^{(k)}(n)
\end{aligned} \tag{3}$$

In equation (2), 5 symbols have been plotted for the pilot channel and 10 for the traffic channel. It is assumed that the temporal length of the physical channel is L chips.

IV. A family of space-time receivers

With this model in mind and modeling the interference-plus-noise as spatially and temporally correlated Gaussian noise, it is possible to formulate the likelihood function which, appropiately minimized, gives the detector for the unknown symbols:

$$J = (\mathbf{y}-\mathbf{Hd})^H \mathbf{R}_w^{-1}(\mathbf{y}-\mathbf{Hd}) = \mathbf{y}^H \mathbf{R}_w^{-1}\mathbf{y} - 2\,\mathrm{Re}\{\mathbf{y}^H \mathbf{R}_w^{-1}\mathbf{Hd}\} + \mathbf{d}^H \mathbf{H}^H \mathbf{R}_w^{-1}\mathbf{Hd} \tag{4}$$

Some assumptions are possible so as to simplify the receiver:

A.1. The correlation matrix of the noise-plus-interference can be decomposed as the Kronecker product of a spatial correlation matrix and a time correlation matrix:

$$\mathbf{R}_w = \mathbf{R}_{w,s} \otimes \mathbf{R}_{w,t} \tag{5}$$

which agrees with the most recent channel models [Pedersen], in which the time and angular spreads are shown to be independent phenomena.

A.2. The physical channel spread (L chips) is much shorter than the length of the spreading code (which is the case when designing a DS/CDMA system), so the matrix $\mathbf{H}^H \mathbf{R}_w^{-1}\mathbf{H}$ is almost diagonal and the last term in (4) can be neglected in its minimization. In fact, this is one of the reasons why high bit-rate users cannot be allocated in rural or hilly environments, where delay spreads are usually long compared to the length of the spreading codes.

Therefore only the middle term remains in (4) and it constitutes a sufficient statistics of the problem. Its maximization leads to the well known Rake receiver when both space and time correlation matrices are assumed white:

$$\begin{aligned} \hat{\mathbf{d}} &= \arg max\, \mathrm{Re}\{\Im\} \\ \Im &= \mathbf{y}^H \mathbf{R}_w^{-1}\mathbf{Hd} = \mathbf{y}^H \left(\mathbf{R}_{w,s}^{-1} \otimes \mathbf{R}_{w,t}^{-1}\right)\mathbf{Hd} = \\ &= \mathbf{y}^H \left(\mathbf{R}_{w,s}^{-1/2} \otimes \mathbf{R}_{w,t}^{-1/2}\right)\left(\mathbf{R}_{w,s}^{-1/2} \otimes \mathbf{R}_{w,t}^{-1/2}\right)^H \mathbf{Hd} = \mathbf{y}_B^H \mathbf{H}_B \mathbf{d} \end{aligned} \tag{6}$$

The introduction of the correlation matrix of $\mathbf{w}$ implies a prewhitening of both the signal vector $\mathbf{y}$ (which is noted with $\mathbf{y}_B$) and the desired user channel matrix $\mathbf{H}$ (which is noted with $\mathbf{H}_B$). This operation can be done separately in time and space (note that $\mathbf{R}_{w,s}^{-1/2}$ apply only on the spatial components of $\mathbf{y}$ and $\mathbf{R}_{w,t}^{-1/2}$ apply on the temporal components). Figure 3 is a representation of the operations performed by this receiver.

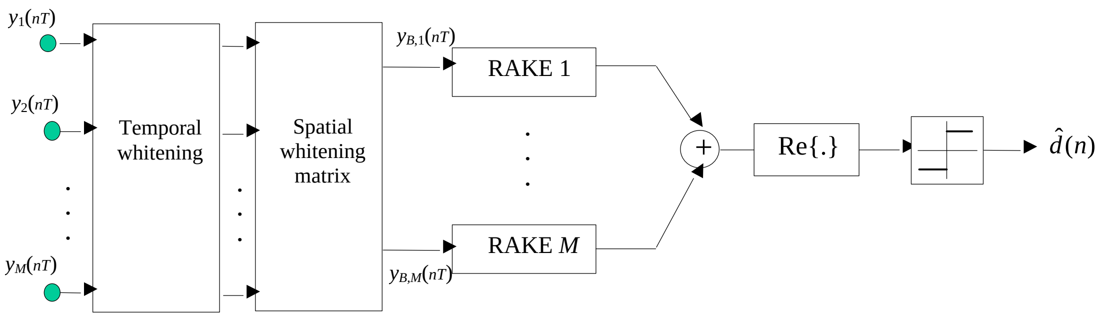


Figure 3. The receiver in equation (6) with temporal and spatial prewhitening matrices. Accordingly, the spatio-temporal channel H has to be temporally and spatially whitened before being used in the rake combiners.

Of course this receiver could be fully implemented by using sample estimates of both correlation matrices, but it is usually the case that the complexity of the resulting structures does not justify the improvement obtained with simplified versions. These different receivers can be formulated from equation (6) by doing certain approximations on the correlation matrices.

A. *Temporal correlation matrix*

T.1. Temporally white interference. It is assumed usually and is a reasonable assumption if the number of interferent users can be considered high.

T.2. *p*th order Markov model for the temporal correlation. This assumption is not usually used but it works well for a low number of users in a low noise scenario. The simplicity of this model is reflected in the structure of the temporal matrix for the first order case, which takes the form:

$$\mathbf{R}_{w,t} = \sigma^2 \begin{bmatrix} 1 & \rho & \cdots & \rho^{N-1} \\ \rho & 1 & \cdots & \rho^{N-2} \\ \vdots & \vdots & \ddots & \vdots \\ \rho^{N-1} & \rho^{N-2} & \cdots & 1 \end{bmatrix} \tag{7}$$

this matrix, and those obtained for higher order models, have in fact a closed expression for its inverse [Kay] which could be used in (6) and prevents from matrix inversion. However, it is much more interesting and practical to recognize that the temporal whitening role of $\mathbf{R}_{w,t}^{-1/2}$ will be done exactly by a *p*th order FIR filter. Of course, high orders of the model, imply long FIR filters which introduce additional interchip interference and, as a consequence, reduces the validity of assumption A.2. Care should be taken to use short lengths compared to the delay spread of the channel impulse response.

B. Spatial correlation matrix approximations

S.1. Spatially white interference. This assumption is realistic only in the case of a high number of interferers or in a highly angular dispersive scenario. Then, the receiver becomes the well-known VRAKE [VanEtten].

S.2. Reduced rank approximation:

$$\mathbf{R}_{w,s}^{-1} \cong \mathbf{B}\mathrm{S}^{-1}\mathbf{B}^H \text{, where } \mathbf{B} \in \mathbb{C}^{M\times R} \text{ with } R<M \tag{8}$$

The spatial correlation matrix is now reduced to a number of is components and, if temporal whiteness for the interference is assumed, the overall receiver operation can be written as:

$$\Im = \sum_{i=1}^{R} \frac{1}{\sigma_i} \mathbf{y}^H (\mathbf{b}_i \otimes \mathbf{I})(\mathbf{b}_i^H \otimes \mathbf{I})\mathbf{Hd} \tag{9}$$

The eigenvector associated to the beamformer *i* gives a measure of the reliability of the information conveyed by the branch *i* of the combiner.

V. Spatial front-ends

Some approximate spatial receivers will be developed in the sequel. No assumptions are made on temporal correlation matrix, thanks to the spatial-temporal uncoupling of the problems stated in equation (5).

A. Noise-plus-interference matrix inversion (NIMI) receiver

This receiver is based on the reduced rank approximation of the inverse of the spatial correlation matrix given by equation (8). It is illustrative to interpret equation (9) as a coherent combining (maximum ratio combining) of the outputs of *R* beamformers (see figure 4). The nature of each is easily seen from a simple case: assume the case of $P<M$ point interferers. If vectors $\mathbf{b}_i$ are taken as the noise eigenvectors of $\mathbf{R}_{w,s}$ each one acts as a spatial interference canceller. Seen in this way, different interference cancellers can be estimated according to different criteria, as presented below.

B. Spatial reference (SR) receiver

An alternate way to determine the vectors in (9) is to compute *R* spatial reference beamformers which cancel the interferers, each associated to an incoming direction of the desired signal. This approach relies on an angular reference for the user of interest [Widrow]:

$$\min_{\mathbf{b}_i} \mathbf{b}_i^H \mathbf{R}_{y,s} \mathbf{b}_i \qquad \textit{subject to} \quad \mathbf{b}_i^H \mathbf{p}_i = 1 \qquad i = 1,\ldots,R \tag{10}$$

Note that the number of significant signal components $\mathbf{p}_i$ will determine R, the degree of the approximation in (9) and hence the number of rake combiners in figure 4. Each combiner will be given by the classical expression:

$$\mathbf{b}_i = \delta_i \mathbf{R}_{y,s}^{-1}\mathbf{p}_i \qquad \delta_i = \left(\mathbf{p}_i^H \mathbf{R}_{y,s}^{-1}\mathbf{p}_i\right)^{-1} \tag{11}$$

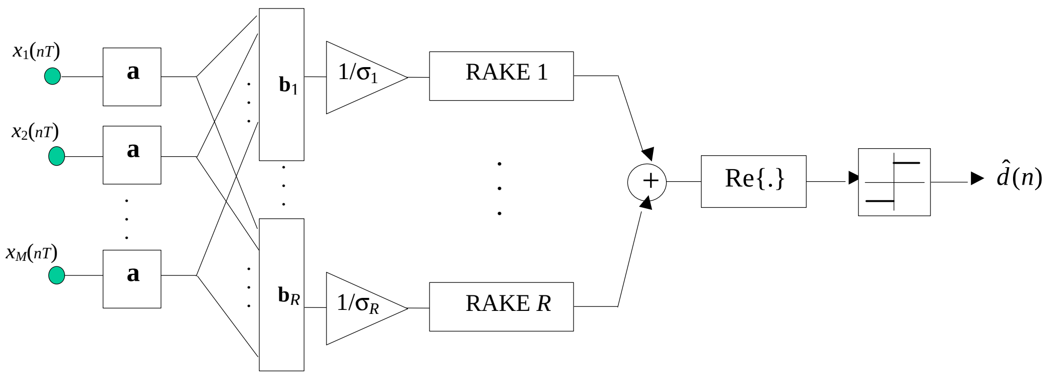


Figure 4. The receiver in equation (6) with temporal prewhitening (given by the FIR filters affecting equally to each branch) and spatial prewhitening and reduced tank approximation (M>R). The gains at the output of each beamformer are given by the associated eigenvalues in equation (8).

A possible method to determine the vectors $\mathbf{p}_i$ is as follows. Let us first redefine the signal model of equation (1) as:

$$\overline{\mathbf{Y}} = unvec_{(N-L+1)\times M}(\mathbf{y}) = \mathbf{D}\overline{\mathbf{H}} + \mathbf{W} \tag{12}$$

where matrix $\mathbf{D}$ is a Toeplitz matrix built at chip time from the QPSK complex spreaded and scrambled symbols:

$$\mathbf{D} = \begin{bmatrix} d(L-1) & d(L-2) & \cdots & d(0) \\ d(L) & d(L-1) & \cdots & d(1) \\ \vdots & \vdots & \ddots & \vdots \\ d(N-1) & d(N-2) & \cdots & d(N-L-1) \end{bmatrix} \in \mathbb{C}^{(N-L+1)\times L} \tag{13}$$

where N stands for the number of chips in the pilot, L is the length of the estimated physical channel, and all terms $d(n)$ belong to the set {-1-j, -1+j, 1+j, 1-j}. $\overline{\mathbf{H}}$ contains the response of the physical propagation channel at chip time for all sensors (note the difference with matrix $\mathbf{H}$ in equation (2)):

$$\overline{\mathbf{H}} = [\mathbf{h}_1 \;\; \mathbf{h}_2 \;\; \cdots \;\; \mathbf{h}_M] \in \mathbb{C}^{(L+1)\times M} \qquad \mathbf{h}_i^T = [h_i(0) \;\; h_i(1) \;\; \cdots \;\; h_i(L-1)] \tag{14}$$

The estimation of the channel matrix can be performed using least squares. Then the vectors $\mathbf{p}_i$ may be obtained as the principal eigenvectors of $\overline{\mathbf{H}}^H\overline{\mathbf{H}}$, or, alternatively, as those eigenvectors pointing to the angular directions in which the SINR is higher. This later condition means choosing the eigenvectors $\mathbf{p}_k$ of $\overline{\mathbf{H}}^H\overline{\mathbf{H}}$ that maximize the signal to signal-plus-noise-plus-interference ratio:

$$SSINR_k = \frac{\mathbf{p}_k^H \overline{\mathbf{H}}^H \overline{\mathbf{H}} \mathbf{p}_k}{\mathbf{p}_k^H \mathbf{R}_{y,s} \mathbf{p}_k} = \frac{\lambda_k}{\mathbf{p}_k^H \mathbf{R}_{y,s} \mathbf{p}_k} \tag{15}$$

In practice this choice proves to be the most effective. The last terms to be defined are the reliability factors σ_k, that is, the noise-plus-interference levels at the output of each beamformer. Since the signal power is equal in all branches thanks to the restriction in (10), it is reasonable (although not strict) to consider the reliability factor to be the total signal power at the output of each beamformer. This term turns out to be δ_i computed in (11).

As a final comment, note that this approach applied to equation (9) is equivalent to consider that the noise-plus-interference correlation matrix is substituted by the expression:

$$\mathbf{R}_{w,s}^{-1} \approx \mathbf{R}_{y,s}^{-1}\mathbf{P}\mathbf{P}^H\mathbf{R}_{y,s}^{-1} \qquad \mathbf{P} = [\mathbf{p}_1 \quad \mathbf{p}_2 \quad \cdots \quad \mathbf{p}_R] \tag{16}$$

C. Matched desired impulse response (MDIR) receiver

A third way to build a spatial receiver is to obtain a combiner **b** that maximizes the SINR at its output. This corresponds to the MDIR approach developed in [Lagunas]:

$$\min_{\mathbf{b}} \mathbf{b}^H\mathbf{R}_{y,s}\mathbf{b} \qquad \textit{subject to} \quad \mathbf{b}^H\overline{\mathbf{H}}^H\mathbf{D}^H\mathbf{D}\overline{\mathbf{H}}\mathbf{b} = 1 \tag{17}$$

It is shown there that the choice of the beamformer is obtained as the eigenvector **b** of the equation:

$$\mathbf{R}_{y,s}\mathbf{b} = \lambda\overline{\mathbf{H}}^H\mathbf{D}^H\mathbf{D}\overline{\mathbf{H}}\mathbf{b} \tag{18}$$

associated to the minimum eigenvalue λ. Being fixed to 1 the signal power at the output of the beamformer **b**, λ takes the value of the inverse of the signal-plus-noise-plus-interference power. Note however that there are M eigenvectors given by (18) and each yield a different signal with different quality. This diversity can be coherently combined using the rake in figure 4, with the eigenvalues used as reliability factors in the branches of the rake.

From the exposition above, the SR receiver and the MDIR receiver seem look very similar. Note however that in the former case, each beamformer points to a different direction, given by the vectors $\mathbf{p}_i$ while in the MDIR approach each beamformer tends to point to all directions from where signals are incoming. The possibilities of the SR receiver seem better, since there are more free degrees of freedom per beamformer to cancel interferences, although the actual performances may depend greately on the angular and temporal dispersion of the channel and have to be determined experimentally.

The most important difference is that, for figure 4 to be valid, the noise components between branches have to be uncorrelated. It is easy to show that this is the case for the MDIR receiver and the NIMI receiver, but not for the SR.

Theorem. The noises at the output of the beamformers obtained with the MDIR approach are uncorrelated, unless the eigenvalues associated are equal.

Proof. Let us take equation 18 and recognise that the same solution for the eigenvectors can be obtained by substituting $\mathbf{R}_{y,s}$ for $\mathbf{R}_{w,s}$. Now let us extend the equation with all the eigenvectors as:

$$\mathbf{R}_{w,s}\mathbf{B} = \overline{\mathbf{H}}^H\mathbf{D}^H\mathbf{D}\overline{\mathbf{H}}\mathbf{B}\mathbf{S} = \mathbf{R}_{d,s}\mathbf{B}\mathbf{S} \tag{19}$$

By left-multiplying with the conjugate transpose of **B** we obtain on the left hand side of the equation, the correlation matrix of the noises at the output of the different beamformers:

$$\mathbf{B}^H\mathbf{R}_{w,s}\mathbf{B} = \mathbf{B}^H\mathbf{R}_{d,s}\mathbf{B}\mathbf{S}$$

Now, since the left-hand side of the equation is an hermitian matrix and the eigenvalues are real we can write:

$$\mathbf{B}^H\mathbf{R}_{w,s}\mathbf{B} = \mathbf{S}\mathbf{B}^H\mathbf{R}_{d,s}\mathbf{B}$$

With no loss of generality assume that S has a multiple eigenvector σ, so it the product above is commutative it can be written as:

$$\mathbf{S}\mathbf{B}^H\mathbf{R}_{d,s}\mathbf{B} = \begin{bmatrix} \sigma\mathbf{I} & \mathbf{0} \\ \mathbf{0} & \mathbf{S}' \end{bmatrix}\begin{bmatrix} \mathbf{C} & \mathbf{D}^H \\ \mathbf{D} & \mathbf{E} \end{bmatrix} = \begin{bmatrix} \mathbf{C} & \mathbf{D} \\ \mathbf{D}^H & \mathbf{E} \end{bmatrix}\begin{bmatrix} \sigma\mathbf{I} & \mathbf{0} \\ \mathbf{0} & \mathbf{S}' \end{bmatrix} \tag{20}$$

By operating it is easy to see that **D**=**0** and that **E** has to be diagonal. Therefore we can conclude that:

$$\mathbf{B}^H \mathbf{R}_{w,s} \mathbf{B} = \begin{bmatrix} \sigma\mathbf{C} & \mathbf{0} \\ \mathbf{0} & \mathbf{S}'' \end{bmatrix} \text{ where } \mathbf{S}'' \text{ is diagonal.}$$

VI. Estimation issues

A. Spatial correlation matrix estimation

The NIMI receiver assumes the knowledge of the matrix $\mathbf{R}_{w,s}$. Relying on the fact that the noise-plus-interference is time uncorrelated with the desired signal we can derive an expression following equations (12), (13) and (14). Note however, that according to the signal structure of the FDD mode of UMTS, we cannot completely determine matrix **D** beforehand since it contains the known chips of the pilot channel but also the unknown chips of the traffic channel:

$$\mathbf{D} = \beta_1 \mathbf{D}_p + \beta_2 \mathbf{D}_t \tag{21}$$

where β_1 is the weighting factor associated to the pilot (known) chips and β_2 is the one associated to the traffic (unknown) chips, as shown in figure 1. First of all, it is worth mentioning that the channel in equation (14) may be estimated consistently by appliyng a least square estimation:

$$\hat{\overline{\mathbf{H}}} = \beta_1^{-1} (\mathbf{D}_p^H \mathbf{D}_p)^{-1} \mathbf{D}_p^H \overline{\mathbf{Y}} \cong (2\beta_1)^{-1} (N - L + 1)^{-1} \mathbf{D}_p^H \overline{\mathbf{Y}} \tag{22}$$

in which incorrelation between known and unknown chips is assumed. Under these premises, the space correlation matrix of interference and noise can be computed ergodically as:

$$\hat{\mathbf{R}}_{w,s} = \overline{\mathbf{Y}}^H \overline{\mathbf{Y}} - \mu \overline{\mathbf{H}}^H \mathbf{D}^H \mathbf{D} \overline{\mathbf{H}} \cong \overline{\mathbf{Y}}^H \overline{\mathbf{Y}} - 2\mu (\beta_1^2 + \beta_2^2)(N - L + 1) \overline{\mathbf{H}}^H \overline{\mathbf{H}} = \hat{\mathbf{R}}_{y,s} - \mu \hat{\mathbf{R}}_{s,s} \tag{23}$$

where in the last equality we have assumed temporal incorrelation between the in-phase and quadrature components of the scrambled chips, and taken into account the different amplitudes of the pilot and traffic channels. The term μ is included with the following purpose: one of the shortcomings of (23) is the matrix substraction, an operation that may lead to non-positiveness of (23) due to estimation errors. Then, the factor μ can be chosen conveniently so as:

$$\mathbf{z}^H \hat{\mathbf{R}}_{w,s} \mathbf{z} = \mathbf{z}^H \hat{\mathbf{R}}_{y,s} \mathbf{z} - \mu \mathbf{z}^H \hat{\mathbf{R}}_{s,s} \mathbf{z} > 0 \qquad \forall \mathbf{z} \tag{24}$$

If this equation has to be positive definite for every possible vector **z**, then the value of μ has to be smaller than the minimum value of the Rayleigh quotient:

$$\mu < min \left\{ \frac{\mathbf{z}^H \hat{\mathbf{R}}_{y,s} \mathbf{z}}{\mathbf{z}^H \hat{\mathbf{R}}_{s,s} \mathbf{z}} \right\} \tag{25}$$

that is, smaller than the minimum eigenvalue of the matrix pencil $\left(\hat{\mathbf{R}}_{y,s}, \hat{\mathbf{R}}_{s,s}\right)$.

B. Temporal correlation matrix estimation

It was stated in section IV that the temporal uncorrelation operation that is performed in equation (6) can be efficiently done using a temporal whitening filter. The design of this filter is straighforward from the linear prediction theory [Haykin]: the Wiener-Hopf equations yield the coefficients of the *p* coefficients whitening filter, provided we dispose of the first *p*+1 terms of the temporal correlation of the interference. A compact formulation can be obtained if equation (1) is rewritten as:

$$\tilde{\mathbf{Y}} = \left(\mathbf{I}_M \otimes \mathbf{D}\right) \tilde{\mathbf{H}} + \mathbf{W} \tag{26}$$

where matrix **Y** is defined in the following way:

$$\tilde{\mathbf{Y}} = \begin{bmatrix} \tilde{\mathbf{Y}}_1 \\ \hdashline \tilde{\mathbf{Y}}_2 \\ \hdashline \vdots \\ \hdashline \tilde{\mathbf{Y}}_M \end{bmatrix} \qquad \tilde{\mathbf{Y}}_i = \begin{bmatrix} y_i(p) & y_i(p-1) & \cdots & y_i(0) \\ y_i(p+1) & y_i(p) & \cdots & y_i(1) \\ \vdots & \vdots & \ddots & \vdots \\ y_i(N-1) & y_i(N-2) & \cdots & y_i(N-p-1) \end{bmatrix} \in \mathbb{C}^{(N-p)\times(p+1)} \tag{27}$$

matrix **D** is defined as in equation (13) (using the symbols at chip time) except for the dimensions:

$$\mathbf{D} = \begin{bmatrix} d(p) & d(p-1) & \cdots & d(-L+1) \\ d(p+1) & d(p) & \cdots & d(-L+2) \\ \vdots & \vdots & \ddots & \vdots \\ d(N-1) & d(N-2) & \cdots & d(N-L-p) \end{bmatrix} \in \mathbb{C}^{(N-p)\times(p+L)} \tag{28}$$

and $\tilde{\mathbf{H}}$ contains the estimated channel at chip time for all sensors:

$$\tilde{\mathbf{H}} = \begin{bmatrix} \tilde{\mathbf{H}}_1 \\ \hdashline \tilde{\mathbf{H}}_2 \\ \hdashline \vdots \\ \hdashline \tilde{\mathbf{H}}_M \end{bmatrix} \qquad \tilde{\mathbf{H}}_i = \begin{array}{c} \\ 1 \\ L \\ \\ \\ \end{array} \overbrace{\begin{bmatrix} \mathbf{h}_i & & \\ & \ddots & \\ & & \mathbf{h}_i \end{bmatrix}}^{p} \in \mathbb{C}^{(L+p-1)\times p} \tag{29}$$

Assuming time incorrelation between the desired user and the interference, as well as incorrelation between the traffic and pilot chips, the time correlation matrix of the interference results in:

$$\hat{\mathbf{R}}_{w,t} = \tilde{\mathbf{Y}}^H \tilde{\mathbf{Y}} - \tilde{\mathbf{H}}^H \left(\mathbf{I}_M \otimes \mathbf{D}^H\right)\left(\mathbf{I}_M \otimes \mathbf{D}\right)\tilde{\mathbf{H}} \cong \mathbf{Y}^H \mathbf{Y} - 2(1+\beta^2)(N-p)\tilde{\mathbf{H}}^H \tilde{\mathbf{H}} \tag{30}$$

From this expression it is possible to recover the coefficients of the temporal whitening filters using Levinson recursion. Note that the regularization device proposed above for the spatial correlation matrix can also be used here.

VII. Experimental performance evaluation

A. Propagation channel model

In order to evaluate the receiver in a realistic mobile scenario, we have carried out simulations based on a Gaussian stationary uncorrelated hypothesis for the channel, assuming independence between angular and Doppler spread, as it has been experienced from measurements taken in downtown Stockholm in the 1,8 GHz band [Pedersen]. There, it is empirically shown that azimuth spectrum follows a Laplacian law, along with Gaussian distribution for the directions of arrival (f) around the mean angular position of the user. The angular spread (that is the standard deviation of the Gaussian, s_f) is taken 8º. The number of rays impinging the array is fitted as a Poisson random variable of mean 25. An exponential law is found in [Pedersen] for the power delay spread, but our simulations will be based on the pedestrian and vehicular models for temporal spreading recommended in the SMG2 documents for UTRA. The amplitude associated with each propagation path (a) is a complex Gaussian random variable whose power decreases as the time delay and the angular direction of arrival with respect to the mobile position increase.

A classical Clarke's bath-shaped Doppler spectrum is obtained by assuming multiple reflections close around the mobile. The carrier frequency is 2,0 GHz. All sensors have flat spatial response in a sectored area of 120º, and are linearly and uniformly spaced at d/λ=0,5. All plots shown in the simulations below are representations of the performance of the link level which can be used later, through convenient mapping, to obtain FER (frame erasure ratio) when considering channel coding or other system level features [Hämäläinen].

B. Simulations

A set of simulations has been performed using up to 9 users of spreading factor 16. All users are assumed to have controlled transmitted power with no error and the propagation channel is the pedestrian, with mobiles moving around at 3 km/h. A diferent number of receivers has been tested: MDIR, SR with different number of eigenvectors, VRAKE and NIMI and the probability of error plotted in figure 5. In all cases, the performance was superior to the conventional VRAKE receiver, so substantial gain from the use of spatial beamforming is achieved. It is found that similar performance is obtained with MDIR and SR receivers, showing no improvements when going from 2 to higher number of eigenvectors (although MDIR shows the best performance when a single eigenvector is used). This is verified in figure 6, where the cumulative function of the ratio of the increasing eigenvalues to the maximum eigenvalue of the MDIR receiver are depicted. It is clear that the second eigenvalue (that is, the SNIR associated to the output of the second eigenvector) is always significant, although it decreases slightly as the number of active users increase. Note that the third eivenvalue is only significant for a low number of users, so it can be discarded.

VIII. Conclusions

Different space-time processors have been presented which boost the power of spatial cancelling and coherent rake combining. They have been tested in a realistic scenario, using UMTS' FDD mode and up-to-date models for spatio-temporal propagation channels. Results show a significant improvement in the probability of error with respect to conventional approaches, that is, only spatial beamforming or only VRAKE combining. Further work is intended to temporally track the estimated channel and correlation matrices parameters so as to reduce bit errors when high speed scenarios are found.

IX. References

[ETSI-UTRA] "Submission of Proposed Radio Transmission Technologies: the ETSI UMTS Terrestrial Radio Access (UTRA) ITU-R RTT Candidate Submission", ETSI SMG2. Date of submission: 29/1/1998. Available at the ITU WWW http://www.itu.ch/imt/

[Hämäläinen] S. Hämäläinen, P. Slanina, M. Hartman, A. Lappeteläinen, H. Holma, O. Salonaho, "A Novel Interface between Link and System Level Simulations", *Proc. of the ACTS Mobile Telecommunications Summit*, Aalborg, Denmark, October 1997, pp. 599-604.

[Haykin] S. Haykin, *Adaptive Filter Theory*, Prentice Hall, 1996.

[Jung] P. Jung, J. Blanz, "Joint detection with coherent receiver antenna diversity in CDMA mobile radio systems", *IEEE Trans. On Vehicular Technology*, vol. 44, no.2, Feb. 1995.

[Kay] S. Kay, *Modern Spectral Analysis*, Prentice-Hall.

[Lagunas] M.A.Lagunas, J. Vidal, A.I. Perez, "Joint beamforming and Viterbi equalizers in wireless communications", *Proc. 31st Asilomar Conf. On Signals, Systems and Computers*, Nov. 1997.

[Mestre] X. Mestre, J. R. Fonollosa, "Algorithms for Flexible Multi-Standard Array Processing: Part 3", Deliverable D711, AC347/UPC/A72/PI/I007/b1, ACTS 0347 SUNBEAM project.

[Pedersen] K. Pedersen, P. Mogensen, B. Fleury, "A Stochastic Model of the Temporal ans Azimuthal Dispersion seen at the Base Station in Outdoor Propagation Environments", *submitted to IEEE Trans. on Vehicular Technology*.

[Rupf] M. Rupf, F. Tarköy, J. L. Massey, "User-Separating Demodulation for Code-Division Multiple Access Systems", *IEEE Journ. Select. Areas in Comm.*, June 1994, pp. 786-795.

[VanEtten] W. Van Etten, "Maximum Likelihood Receiver for Multiple Channel Transmission Systems", *IEEE Trans. on Communications,* Feb. 1976, pp. 276-283.

[Verdú] S. Verdú, *Multiuser Detection*, Prentice-Hall, 1998.

[Widrow] B. Widrow, P.E. Mantey, L.J. Griffiths, B.B. Goode, "Adaptive Antenna Systems", Proc. IEEE, vol. 55, pp. 2143-2159, 1967.

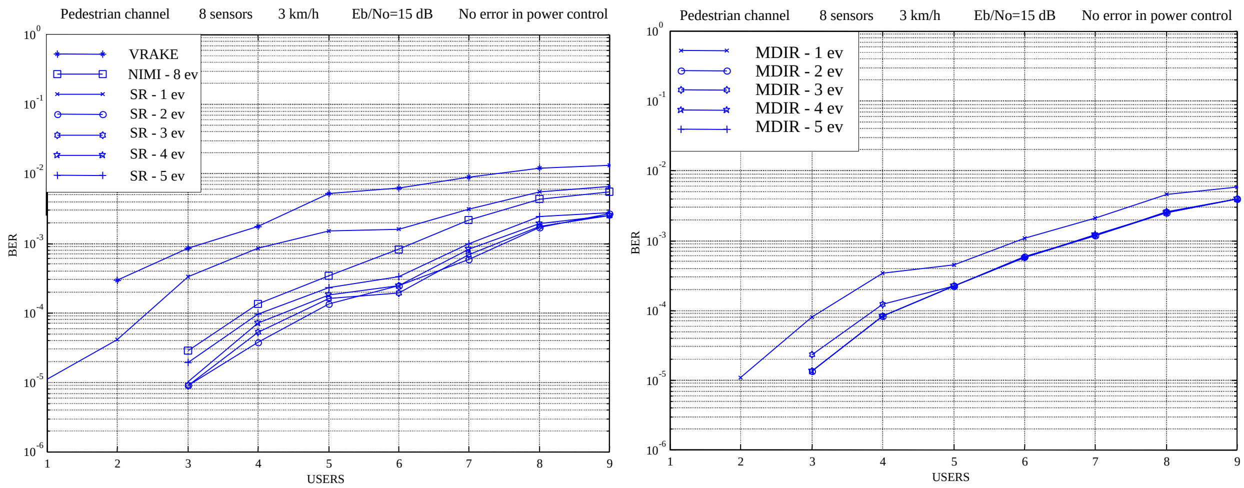


Figure 5. Probability of error for a different number of active users, all transmitting controlled power, for the different receivers and different number of combiners.

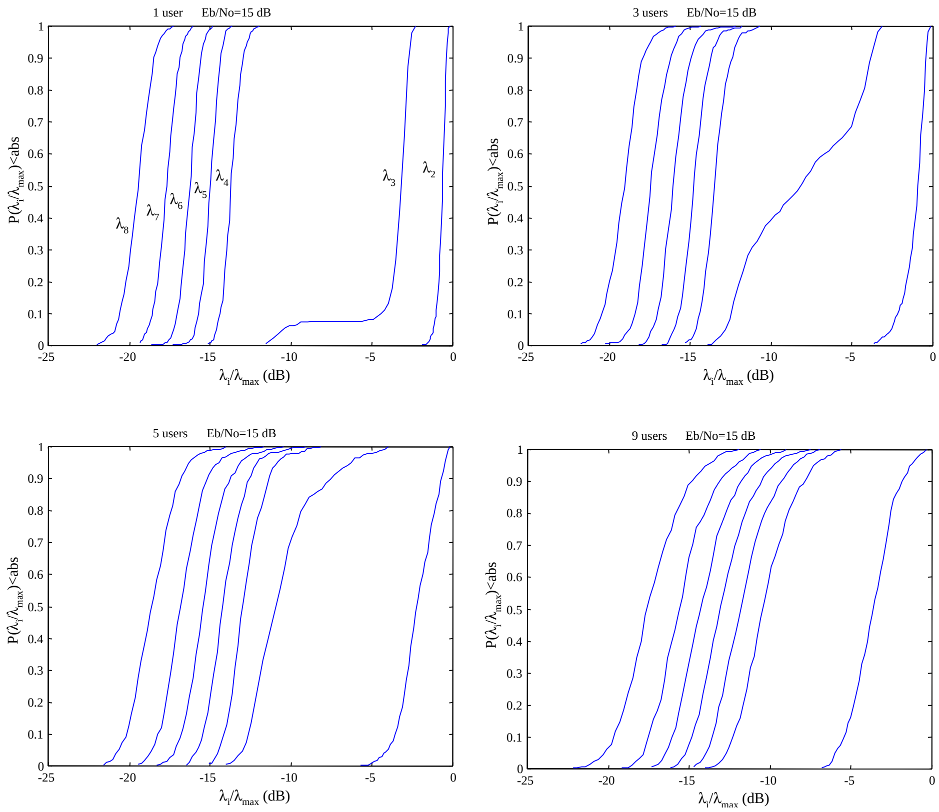


Figure 6. Cumulative functions of the ratio of the MDIR eigenvalues to the maximum eigenvalue. Note that the second eigenvalue is always significant, while the third eigenvalue becomes less significant as the number of active users increase.